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Micro-Stripline Antennas

In micro stripline circuits discontinuities occur which may be defined as all deviations from the uniform stripline structure such as bends, kinks, branches, impedance transitions, open and short circuits and losses through radiation. Radiation losses increase with frequency, particularly when the geometrical dimensions of the line approach those of an electrical wavelength. This radiation is encouraged in stripline antennas by a suitable arrangement of these discontinuities. Particularly good radiation occurs from power resonators and surface resonators whose electrical dimensions are λ / 2 or multiples thereof. This fact is utilised in other types of antenna.

Micro-stripline antennas are therefore radiating surfaces on a thin dielectric substrate with a conducting ground plane. The main radiation axis is perpendicular to the plane of the radiator.

1. ADVANTAGES AND DISADVANTAGES

Owing to its planer structure the antenna is manufactured by the etching technique in exactly the same manner as for micro-stripline circuits. The demands upon this technique and the materials, particularly for the low-loss dielectric, are very similar. The advantage of these antennas are their uniform, thin structure and negligible weight.

The diagonal measurements of the base plane must, however, be at least a wavelength, and for antenna arrays, considerably larger. The arrangement of many radiation elements into an array results in a greater beam concentration and therefore a higher gain.

Owing to the geometrical dimensional limitations, there has to be a lower frequency restriction of about 300 MHz for these antennas. The upper frequency limit is set by rising conductor losses and losses in the dielectric and is around 30 GHz for PTFE dielectric. Surface wave radiations are also intensified with increasing frequency, particularly when the ratio of substrate thickness to wavelength is greater than 0.1.

Since the radiation surface is driven into resonance, these antennas have a narrower working frequency range than a horn radiator. The obtainable bandwidths range from 1 % to 5 % of the mid-band frequency in the 3 cm band.

The efficiency η is also lower than a horn radiator owing to the higher losses in the thin conductor structure. The gain from a well-matched single element is about 6 dB and with a little more (worthwile) trouble with arrays of elements, gains

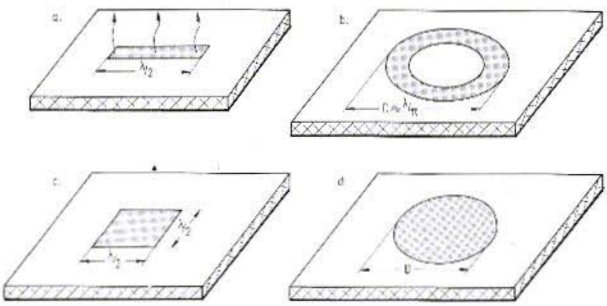

Fig. 1: Surface radiators in micro-stripline technique

of up to 20 dB are possible. The gain, however, is always a few dB lower than the equivalent sized horn antenna would be.

Owing to the negligible power handling capabilities of these antennas, they are not suitable for use as high power radiators. They are almost ideal for receive functions however, as they may be fabricated on the same substrate as say an amplifier with direct low-loss connections – all in the same etching process.

2. FUNCTION

Since the introduction of stripline antennas some 30 years ago a number of technical publications about the subject, and the calculation of fields and radiation conditions, have appeared. For those wishing a more theoretical treatise the reference books (1) and (2) are recommended. They, in turn, provide further references for a deeper study.

In the following article I would like to present a clear description of the micro-stripline antennas together with a few formulae for the approximation of the dimensions.

As already mentioned, conductors radiate particularly powerfully when they complete a $\lambda / 2$ resonate circuit or multiples thereof. Through the nature of the radiating surface and the manner in which they are coupled, they may be categorized into three groups:

The first group encompasses resonant surfaces and conductors and the second are radiating aperatures which to many are known as slot radiators. The third group form travelling wave antennas. The latter comprise a periodic arrangement of discontinuities on a non-radiating feedline which has been properly terminated. As far as the principle function is concerned, travelling wave antennas are similar to surface resonator arrays and will not be considered further.

A few examples of conducting radiating elements are shown in **fig. 1**. A $\lambda / 2$ conductor resonator which produces a linear polarised wave with the E field vector in plane of the antenna is shown in **fig. 1 a**. The ring resonator in **fig. 1 b** represents a closed conductor. The basic oscillitory resonator must have an electrical length of a full wavelength, requiring an average diameter of λ / π. The radiation characteristic and the polarisation, depend upon the way in which it is fed. This applies

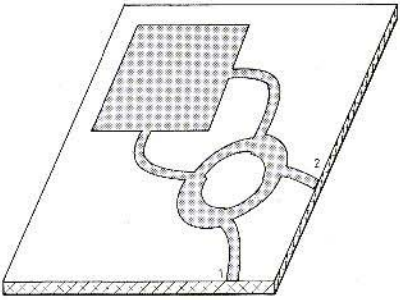

Fig. 2: Circular polarised surface radiator with a 90° hybrid coupler feed in order that the sense of the polarisation may be reversed

also to the surface radiators depicted in 1 c and 1 d as to whether they are linear or circular polarised antennas. The circular polarised radiator must be fed simultaneously on two sides offset physically by 90° by signals having a 90° phase difference between them. The arrangement is shown in **fig. 2**. By the coupling of the surface radiators with a 90° hybrid coupler the polarisation rotational direction may be chosen remotely. If input 1 is fed, and input 2 terminated, the antenna possesses a rigth-handed polarisation and vice-versa a left-handed polarisation is obtained.

The radiating characteristic of the round surface radiator of **fig. 1 d** is dependent upon the type of resonator field which with the aid of cylinder-functions – i. e. approximations to Bessel functions – may be calculated. This cannot be gone into here but references /1/ and /2/ deal with it.

Surface radiator elements may be relatively easily arranged into an array of antennas as **fig. 3** shows. The radiation characteristic and the input impedance of the antenna is determined by the type and phase disposition of the feed. The arrangement in **fig. 3** consists of radiating elements fed in-phase. In the upper diagram $\lambda / 2$ radiating elements are connected by $\lambda / 2$ non-radiating (almost) elements in series so that each radiator leading edge is exactly one wavelength from its neighbour. This means that the antenna is fed in-phase and that the direction of polarisation is in the plane of the antenna.

Fig. 3 lower shows another arrangement of in-phase feeding of radiating elements, the feeder sections are this time, one wavelength long. The polarisation of this arrangement is at right-angles to the feeder section i. e. out of phase by 90° with that of the upper array.

An example of slot radiators is shown in **fig. 4** which depicts two antennas with differing feed arrangements. **Fig. 4 a** shows an arrangement for symmetrical stripline, known as triplate, in which the radiating slot is coupled by a conductor short-circuit. The current is greatest at the short-circuit point and with it the magnetic component of the field which lies at rigth-angles to the plane of the line direction. Since slot antennas have a magnetic field vector which lies in the plane of the antenna, the H field direction of the antenna and that of the short-circuited line are the same so that both slot and line are well coupled.

The arrangement in **4 b** depicts a slot antenna etched in the ground-plane of an unsymmetrical conductor structure and coupled by a micro-stripline which passes over the middle of the slot and at rigth-angles to it. The line is left open-circuited

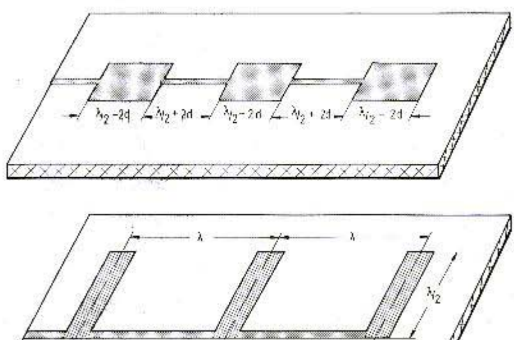

Fig. 3: Antenna arrays consisting of periodically fed radiating elements

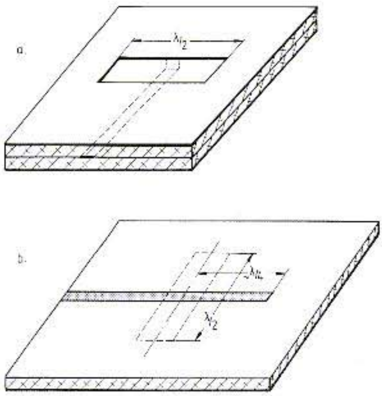

Fig. 4: Examples of slot radiators with H field coupling
a) Triplate configuration with short-circuit coupling
b) Slot radiator in the ground surface with an open-circuit λ / 4 micro-stripline coupling

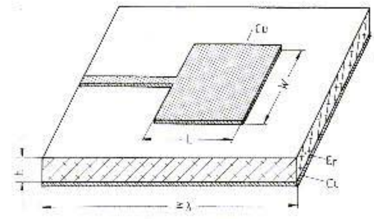

Fig. 5: Rectangular surface radiating elements of length L and width W on a dielectric substrate of thickness h and a complete metal ground screen

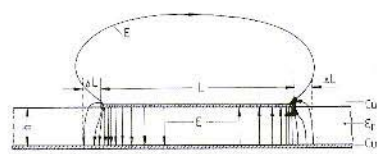

Fig. 6: Cross section through the conductor structure depicting the course and intensity of the electrical field line E

at its end λ / 4 from the middle of the slot. The open-circuit is transformed as a short-circuit in the plane of the slot. The magnetic field component is thereby greatest directly above the slot, which also has its greatest magnetic component at this point, thus providing a good coupling.

The radiation of the slot from the triplate configuration takes place from only one side, the other (ground side) being the reflector. The micro-stripline slot feed arrangement, radiates almost equally in both directions as apposed to the conducting radiators where the metal ground-plane also acts as a radiator, particularly when it has a large area relative to the radiating surface.

3. CALCULATION OF A RECTANGULAR RADIATING ELEMENT

In order to simplify this review, the mathematical details have been omitted but they are available, if required, from ref. /3/.

The surface radiator element shown in **fig. 5** may be regarded as a conductor of length L and width W. Ignore for the moment the narrow feedline and consider it as of length L = λ / 2 i. e. a half-wave resonator.

As the field lines shown in **fig. 6** indicate, at the open-circuit ends, the field is distorted in respect to the ideal field structure along the conductor. This distortion has the effect of elongating the conductor L by the amount △ L. This short conductor length is effectively capacitive with a capacitance C. At the same time, the field lines occur in the free-space from end to end of the conductor and the end may be regarded as a radiating slot of length △ L and of width W. The radiating energy can be considered to be dissipated through a radiation resistance R so that the surface radiator element may be considered as the equivalent circuit shown in **fig. 7**. It can be seen that the resonant frequency f_0 of the resonator is mainly determined by length L and the effective elongation △ L:

(1) $$L = 0{,}5\,\lambda - 2\,\triangle L$$

Since the wave distribution on the conductor is

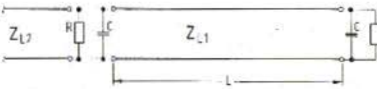

Fig. 7: Equivalent circuit of a radiating element from fig. 5 as a feed-line of characteristic impedance Z_{L1} and the equivalent elements R and C of the open end together with the characteristic impedance Z_{L2}.

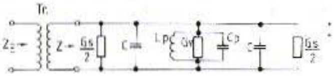

Fig. 8: Equivalent circuit of a micro-stripline radiator in the vicinity of the resonant frequency

Fig. 9: Transformer coupling of a radiating element with a coaxial cable through the ground plane (from /1/)

delayed by the effect of the dielectric, the wavelength is given by

(2) $$\lambda = \lambda_o / \sqrt{\varepsilon_{reff}}$$

where λ_o is the free-space wavelength from

(3) $$\lambda_o = c_o / f_o$$

where c_o = velocity of EM wave in space.

The effective relative permittivity ε_{reff} is dependent upon the conductor width W and the relative permittivity of the substrate. It is given below and in equation /4/ ref. /3/.

(4) $$\varepsilon_{reff} = (\varepsilon_r + 1) / [2 + (\varepsilon_r - 1) \sqrt{4 + 48\,h / W}]$$

The effective elongation $\triangle$ L is calculated as described in ref. /3/ equation /7/.

If the heigth of the substrate $h << \lambda_o$ then $\triangle$ L is approximately: –

(5) $$\triangle L = h / 2$$

The radiation resistance R depends upon the size of the radiating slot and therefore upon the width W of the conductor in relationship to the wavelength.

According to /5/ for the conductor width $W < \lambda / 2$ of the radiation resistance

(6a) $$R = 180\,\Omega / \sqrt{\varepsilon_{reff}} \cdot (\lambda / W)^2$$

and for the conductor width $W > 1.5\,\lambda$ with

(6b) $$R = 240\,\Omega / \sqrt{\varepsilon_{reff}} \cdot (\lambda / W)$$

Since the radiation resistance is effective on both sides of the conductor, the resultant value R_S for one element is:

(7) $$R_s = R / 2$$

The width W of the conductor is calculated as in ref. /3/ which also gives the characteristic impedance Z_{L1} of the resonator element.

For coupling the antenna to the feedline a knowledge of the total impedance Z is necessary and regarding the equivalent circuit of **fig. 8**, the total value of the conductance. This comprises the radiation conductance G_s, a loss conductance G_v, and a susceptance of the equivalent elements of the conductor L_L, C_P and 2 C.

The matching transformer is either accomplished in the feed-line or through coupling into the radiating elements within the length L (**fig. 9**) so that the resonator line itself is used as a transformation element.

For the case of resonance at f_o, the reactive components should fall to zero and the transformed quantities of the radiation conductance and the loss conductance will form the desired input impedance Z_E. In order to have a good antenna efficiency, it is necessary that the loss conductance is very small in relationship to the radiation conductance. Usable micro-stripline antennas are therefore only built with very low-loss dielectrics. The normal micro-stripline materials such as RT/Duroid, Di-Clad or A_2O_3 ceramic, can be considered as practically loss-free up to the 3 cm band. PCB material such as G 10 is unusable above 1 GHz.

Connecting several radiating elements together to form an array, as in **fig. 3**, increases the radia-

tion conductance but also, unfortunately, the loss conductance as well. Additional losses in the connecting lines have the effect of decreasing the ratio of radiation-to-loss conductance. The efficiency of large micro-stripline antenna arrays is so poor that the maximum available gain is limited to 20 dB /6/, /7/ and /8/.

An important criterion of an antenna is its bandwidth, that is, the frequency range in which it is usable. In /1/ the bandwidth for a single radiator was calculated in detail. A knowledge of the material losses is required beforehand, but this can be very tedious for arrays having several radiating elements and coupling networks. A measurement method of determining the bandwidth is very simple and consists of measuring the input impedance versus frequency by means of say, the input VSWR.

The bandwidth Δf will then be defined as the band over which the antenna is suitably matched, i. e. a VSWR of less than say 2 : 1. This can also be expressed as the frequency band over which the return loss is smaller than − 10 dB rel. ref. freq.

Before the measurement is carried out, however, it must be determined that the antenna is beaming in the desired direction in this band of frequencies. It is particularly the case with antenna arrays, it is possible to find that within the usable band and at a frequency at which the VSWR is low, that the radiation beam has slewed off the desired direction or that a side-lobe possesses a greater proportion of the input power. Another possibility, is that the coupling network is matched unintentionally with the loss conductance at this frequency. The radiating elements are then not in resonance and the radiation conductance is negligible. The antenna is then acting as a terminating resistance!

4. CALCULATION AND CONSTRUCTION OF A MICRO-STRIPLINE ANTENNA FOR THE 3 cm BAND

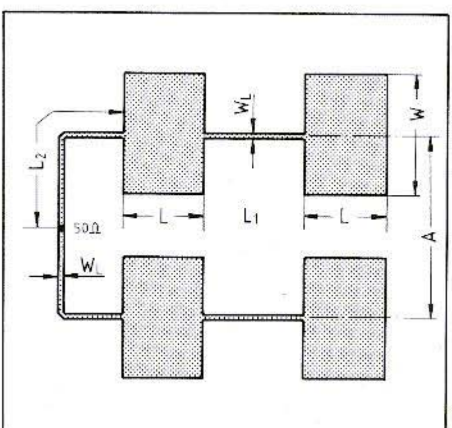

Fig. 10: Antenna array consisting of 2 x 2 rectangular micro-stripline resonators for the 3 cm band

In order to demonstrate the applicability of the formulae, an antenna array consisting of 2 x 2 radiating elements was designed and measured. The array arrangement is shown in **fig. 10**. The array consists of two lines of $\lambda / 2$ radiators, each line having two elements. The polarisation lies in the plane of the feedline L_1.

The radiator length L is determined by the resonant frequency f_o = 10.35 GHz and the width W determines the radiation resistance so that an input impedance of 50 Ω results. In order to have a feed impedance of 50 Ω, each element must have a radiation resistance R_S = 200 Ω on the condition that each element is sufficiently decoupled from the others. The coupling line to the radiator is $L_2 = \lambda$ and has a characteristic impedance of 100 Ω i. e. the impedance of two paralleled radiators.

The construction utilises a teflon substrate RT/Duroid 5880 with a substrate thickness h = 0.5

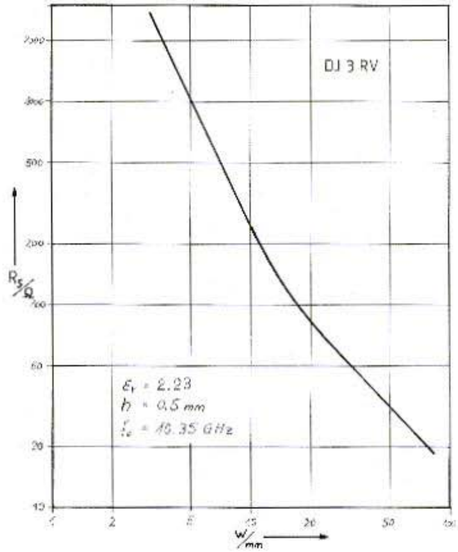

Fig. 11: Radiation resistance R_S of a λ / 2 surface radiator as a function of the conductor width W

mm and 17.5 μm copper film on both sides. The relative permittivity ε_r = 2.23 at 10.35 GHz.

With these material data, a calculation was made for the dimensions L and W of a radiating element as in **fig. 5** at 10.35 GHz and a radiation resistance R_S of 200 Ω. As the radiation resistance lay outside the validity of the formulae **6 a** and **6 b**, the diagram of **fig. 11** was developed. It contains the transitional range of the two formulae as an average value. The diagram directly supplies the radiator width W in millimetres for a given radiation resistance R_S. Now, with the width W, the effective relative permittivity is found with equation (4) and with equation (1) the resonator length L.

The following calculation steps use the diagram from /3/ and the antenna geometrical data for **fig. 10** is listed as follows:

Radiator Width

$R_S = 200\ \Omega \rightarrow$ fig. 11 $\rightarrow$ **W = 11 mm**

Radiator Length

$W = 11$ mm $\rightarrow$ with $h = 0.5$ mm $\rightarrow W / h = 22$

$W / h = 22 \rightarrow$ diagram 1 $\rightarrow \varepsilon_{reff} = 2.11$

$f_0 = 10.35$ GHz $\rightarrow$ (2) + (3) $\rightarrow \lambda = 19.94$ mm

with $L = 0.5\,\lambda - 2 \triangle L$

$\triangle$ L determined from /3/, diagram /3/

$w / h = 22 \rightarrow$ diagram 3 $\rightarrow d / h = 0.68$

$\triangle L = 0.34$ mm

L = 9.29 mm

The feedlines L_1 and L_2 are loaded with the characteristic impedance $Z_L = 100\ \Omega$.

Line Length L_1

$Z_L \rightarrow$ diagram 2 $\rightarrow W / h = 0.9$

$W / h = 0.9 \rightarrow h = 0.5$ mm $\rightarrow$ **W_L = 0.45 mm**

$W / h = 0.9 \rightarrow$ diagram 1 $\rightarrow \varepsilon_{reff} = 1.78$

$f_0 = 10.35$ GHz $\rightarrow$ (2) + (3) $\rightarrow \lambda_1 = 21.71$ mm

With $L_1 = 0.5\,\lambda_1 + 2 \triangle L$ for equal phase feeding of both elements.

L_1 = 11.53 mm

The **distance A** is arbitrary chosen with

A = L + L_1 = 20.82 mm

For the **Line Length** $_2$, the elongation $\triangle$ l for the compensatory kink as in /3/ as well as the reference plane displacement d_1, must be taken into account. Here the feed point is, (as opposed to fig. 10) a T-branch with 50 Ω feed lines.

$$L_2 = \lambda_1 + \triangle L - 2 \triangle l + d_1$$

This results in the length L_2 along the edge to the middle of the antenna.

$W / h = 0.9 \rightarrow$ /3, eq. (9)/ $\rightarrow 2 \triangle l = 0.013$ mm

$Z_1 = 100\ \Omega$

$Z_2 = 50\ \Omega \rightarrow$ /3, (10) + (11)/ $\rightarrow d_1 = 0.133$ mm

L_2 = 22.17 mm

b = 0.231 mm from /3/.

5. ANTENNA MEASUREMENTS

The input impedance characteristic with reference to the feed point in **fig. 10** is shown in the

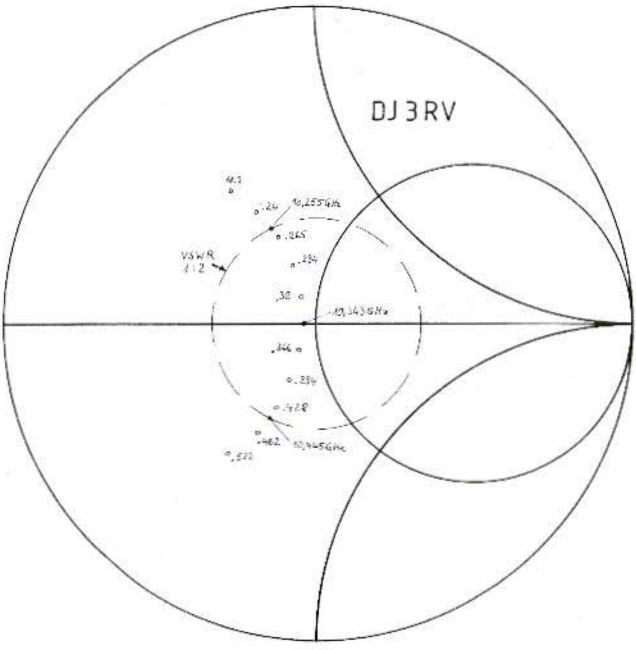

Fig. 12: Feed point input impedance plot on Smith-Chart. The values are for |r| = 0.1; 0.2; 0.3; 0.4 and 0.5 at VSWR 1 : 2 ref. 50 Ω

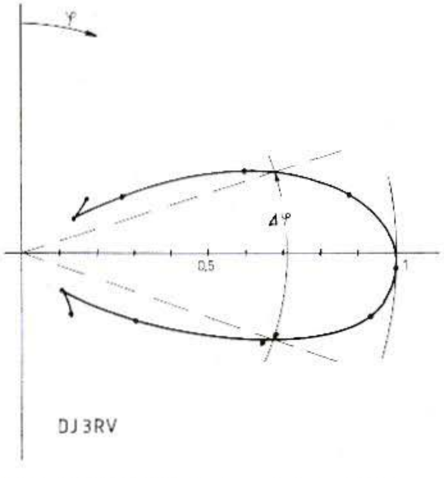

Fig. 13: Directional characteristic in the E-plane (horizontal diagram)

Smith Chart plot in **fig. 12**. The minimal reflection coefficient with

$$r = 0.04\ e^{-180^{\circ}}$$
$$\text{at } f_o = 10.343 \text{ GHz}$$

This corresponds to an input resistance of

$$Z_E = 46\ \Omega$$

A VSWR = 2 : 1 is measured at 10.255 GHz and 10.445 GHz, giving a bandwidth of 190 MHz i. e. 1,84 %.

Fig. 13 shows the polar plot in the E field plane at f_0 = 10.343 GHz. That is the polarisation plane and, normally, the horizontal diagram. It is probably something to do with the feed arrangements that there is a +2° departure from the normal plane symmetry.

The half-power points:

$$\triangle\varphi = +19^{\circ} \text{ to } -17^{\circ} = 36^{\circ}$$

and the first minima occurs at

$$\varphi\, m = +42^{\circ} \text{ to } -34^{\circ} = 76^{\circ}$$

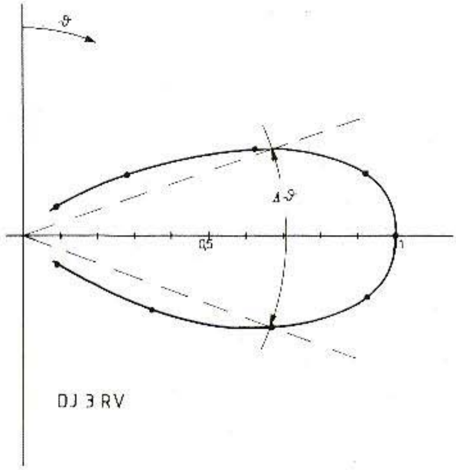

Fig. 14: Directional characteristic in the H-plane (vertical diagram)

Fig. 14 shows the directional characteristic in the H plane at $f_0 = 10.343$ GHz.

The half-power points

$\triangle \vartheta = +20° \text{ to } -19° = 39°$

and the first minima occur

$\vartheta m = +42° \text{ to } -41° = 83°$.

The **Gain** was measured at $G_i = 10.5$ dB_i

The cross-field polarisation (E component measured in H plane) = − 36 dB (only).

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